FREQUENCY COMPENSATION TECHNIQUES FOR AN INTEGRATED OPERATIONAL AMPLIFIER *

by James N. Giles

I INTRODUCTION

Application of negative feedback to a high-gain amplifier gives a circuit with characteristics dependent almost entirely upon the feedback elements. The improvement in gain stability, phase shift, input impedance, output impedance, and linearity is proportional to the amount of feedback. Thus, amplification to any degree of accuracy is possible with sufficient feedback. Large feedback, however, requires that close attention be given to the amplifier open-loop characteristics. Stable circuits must have well-defined open-loop gain and phase responses to frequencies far above the band of interest: 60 db of feedback over a 10-kHz bandwidth, for example, requires controlled open-loop characteristics to over 10 MHz.

The frequency compensation networks required to stabilize a feedback amplifier depend, among other things, on the open-loop gain, the frequency response characteristics, and the impedance at the compensation terminals of the amplifier. Specification of limits for each individual factor which affects the design of the frequency compensation networks will give an unnecessarily restrictive result since variations in these factors correlate to some degree in an integrated circuit. It is better to treat the amplifier as a whole by specifying complete networks which guarantee stability for the full distribution of units.

This paper presents design techniques for frequency compensation which are generally applicable to most integrated operational amplifiers. The Fairchild μA702A is taken as a specific example for detailed discussion; frequency compensation networks are derived which provide stability for practically any circuit using this amplifier. The techniques used may be applied to other amplifiers if the production distribution of certain circuit characteristics are known.

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II DESCRIPTION

Lag Compensation

The closed-loop gain of the simple feedback amplifier shown in Fig. 1 is given by

$$A_v = -\frac{A_{vo}}{1 + \beta A_{vo}} \qquad (1)$$

where

A_v = closed-loop voltage gain

A_{vo} = open-loop voltage gain

β = feedback factor

The quantity $(1 + \beta A_{vo})$ is the amount of feedback and is a direct measure of the improvement obtained in the performance of the amplifier. This factor must be greater than zero for all frequencies if the circuit is to be stable. Hence the magnitude of the loop gain, βA_{vo}, must be less than unity at the frequency where the loop phase shift is 180 degrees. For real β, the maximum amount of feedback is limited to the ratio of the low-frequency value of A_{vo} and the value of A_{vo} at the 180-degree phase shift frequency.

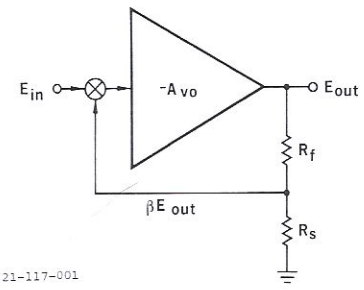

FIG. 1. A simple feedback amplifier.

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The open-loop gain and phase responses of a typical μA 702A amplifier are shown in Fig. 2, including a straight-line approximation to the actual curve. The gain rolls off at the 1- MHz first break frequency with a slope approaching 6 db/octave; a second break occurs at approximately 4 MHz, and a third break at 40 MHz. The 180-degree phase shift frequency is about 14 MHz, so no more than 34 db of feedback may be applied without oscillation. It would not be practical to use this much feedback, however, since it is inevitable that the 180-degree phase shift will be exceeded by minor deviations in the amplifier or external circuitry. Therefore, the limiting phase for maximum feedback should be chosen less than 180 degrees by some definite margin to allow for variations in amplifier and circuit parameters.

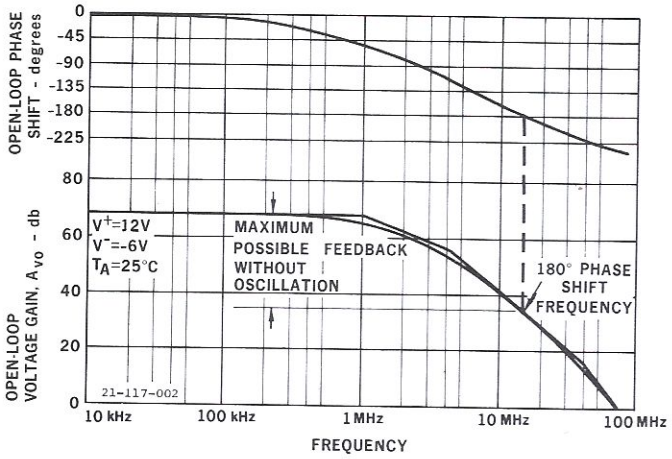


FIG. 2. Typical open-loop gain and phase shift of the μA 702A.

Phase margins less than 90 degrees will cause peaking in the closed-loop response at the frequency where the loop gain equals unity. This is because the magnitude of the denominator of Eq. (1) becomes less than unity due to the phase shift in βA_{VO}. The degree of peaking for various phase margins is shown in Fig. 3. A satisfactory circuit should have a minimum phase margin of about 45 degrees (less than 3-db peaking). Thus only 20 db of feedback can be safely applied to the basic amplifier.

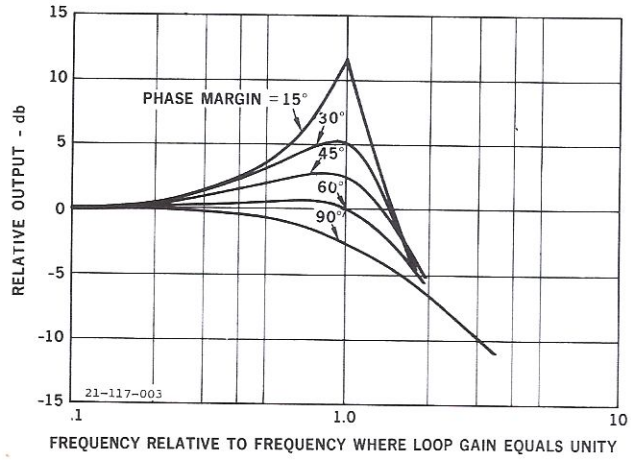


FIG. 3. Peaking in closed-loop response for various phase margins.

Since the amplifier response is limited by internal capacitances, little can be done to increase the amount of feedback without purposely narrowing the open-loop bandwidth. The simplest way of designing a feedback amplifier which is stable with large amounts of feedback is to shunt some signal point in the circuit to ground with a single, relatively large capacitor. If the rolloff due to this capacitor begins at a low enough frequency, the loop gain can be made less than unity before other circuit elements introduce additional phase shift. Since the maximum phase shift associated with a single R-C network is 90 degrees, the circuit cannot become unstable.

A point within the μA 702A has been made available to facilitate this method of bandwidth narrowing. The schematic diagram of the amplifier is shown in Fig. 4; a capacitor connected from the lag compensation terminal to ground attenuates the high frequency signals at the base of Q_7, giving the necessary rolloff in response. Figure 5 shows the open-loop response of an amplifier using this frequency compensation and gives curves for various closed-loop gains.

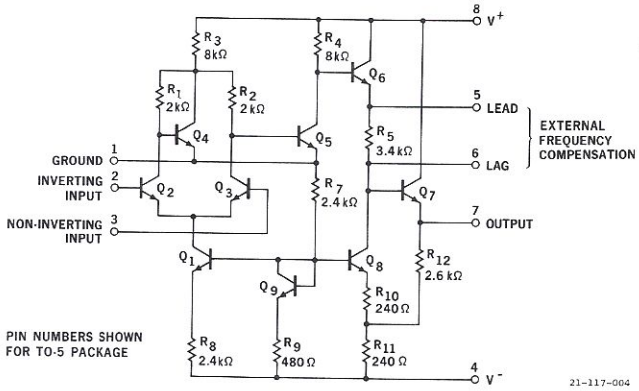


FIG. 4. Schematic diagram of the μA 702A.

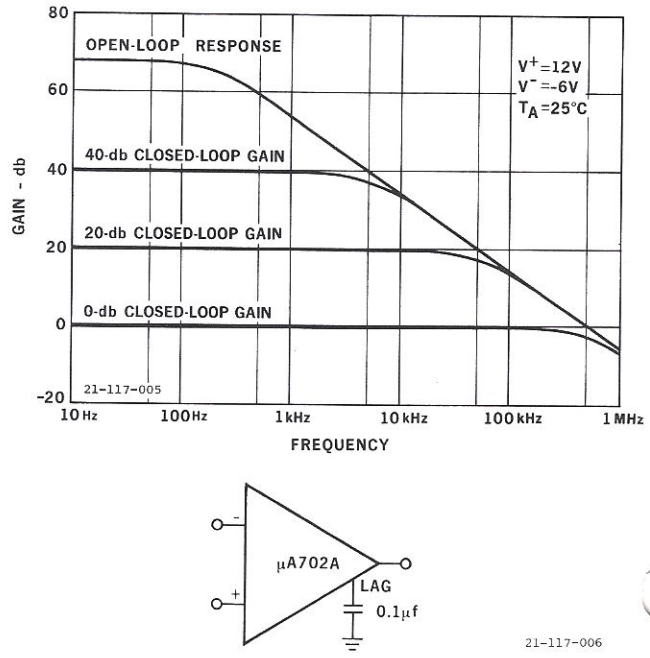


FIG. 5. Frequency response of the μA 702A with simple lag compensation.

Since the closed-loop circuit cannot supply more gain than is available from the amplifier itself, the closed-loop response intersects and follows the open-loop gain curve at high frequencies as in Fig. 5. Clearly, this circuit does not give maximum closed-loop bandwidth. It is ideal for DC applications, however, since frequencies above a few hundred hertz are ordinarily of little interest. The early rolloff in high-frequency gain reduces the broadband noise and makes the circuit less susceptible to instability caused by capacitive loading or stray capacitances in the feedback circuitry.

Maximum feedback over wide bandwidths can be obtained by using the natural rolloff of the amplifier to provide part of the compensation. This technique is illustrated by the circuit and loop-gain response shown in Fig. 6. The series R-C compensation network gives a rolloff in loop gain beginning at some lower frequency and breaking out at the natural rolloff frequency of the amplifier. The amplifier internal rolloff then provides the additional attenuation necessary to bring the loop gain through unity with an adequate phase margin. This scheme gives the widest possible bandwidth for any value of closed-loop gain.

The values of the compensation components depend upon the loop gain, phase margin, and amplifier internal characteristics. The attenuation in loop gain from the compensation network is equal to the ratio of the internal resistance at the compensation terminal and the series compensation resistor; thus

$$A_c = \frac{R_i}{R_1} \tag{2}$$

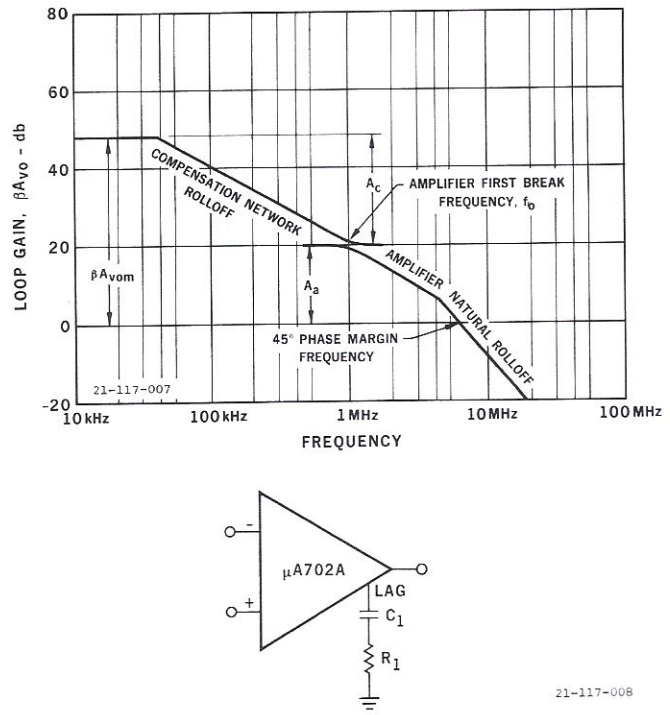


FIG. 6. Compensation for maximum bandwidth with lag network.

where

A_c = attenuation in loop gain from compensation network

R_1 = amplifier internal resistance at the lag compensation terminal

The total attenuation from both compensation and amplifier rolloffs must be equal to the low-frequency loop gain

$$\beta A_{vom} = A_c A_a \tag{3}$$

where

βA_{vom} = magnitude of low-frequency loop gain

A_a = attenuation in loop gain from the amplifier internal rolloff for a given phase margin

Therefore, the required value of compensation resistor is

$$R_1 = R_i \left(\frac{A_a}{\beta A_{vom}}\right) \tag{4}$$

The breakout frequency of the compensation network response should be made equal to the amplifier natural rolloff frequency for maximum feedback bandwidth, hence the value of the compensation capacitor is

$$C_1 = \frac{1}{2\pi f_o R_1} \tag{5}$$

where

f_o = first break frequency of the open-loop response

A generalized feedback circuit for the amplifier is shown in Fig. 7. Single-ended signals may be applied to either input terminal if the resistor at the unused input is returned to

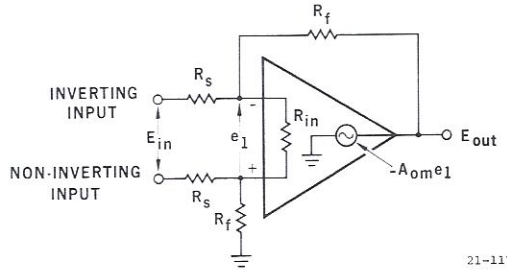


FIG. 7. General feedback circuit.

ground. The total DC resistance to ground from each input should be equal for minimum DC offset at the output. The loop gain for the circuit of Fig. 7 is given by

$$\beta A_{vom} = \left(\frac{1}{1 + \frac{R_f}{R_s}}\right)\left(\frac{A_{om}R_{in}}{R_{in} + 2\frac{R_fR_s}{R_f + R_s}}\right) \quad (6)$$

where

A_{om} = open-loop low-frequency gain measured with zero source resistances

R_{in} = input resistance

Expressions for the closed-loop gain and other circuit characteristics are given in detail by Ref. 1.

Equation (6) shows that the open-loop gain is decreased by circuit resistances which are larger than the amplifier input resistance. This affects the values of the compensation components if the widest possible bandwidth is desired. However, since R_{in} varies greatly with temperature, it is best to design the compensation network assuming that the loop gain is maximum; this ensures stability for all conditions at the cost of some bandwidth with high circuit resistances. This is also important because high circuit resistances make stray capacitances more dominant. Therefore, using Eqs. (4) and (6), and assuming that

$$R_{in} >> 2\frac{R_sR_f}{R_s + R_f}$$

the expression for the compensation resistor becomes

$$R_1 = R_i\frac{A_a}{A_{om}}\left(1 + \frac{R_f}{R_s}\right) \quad (7)$$

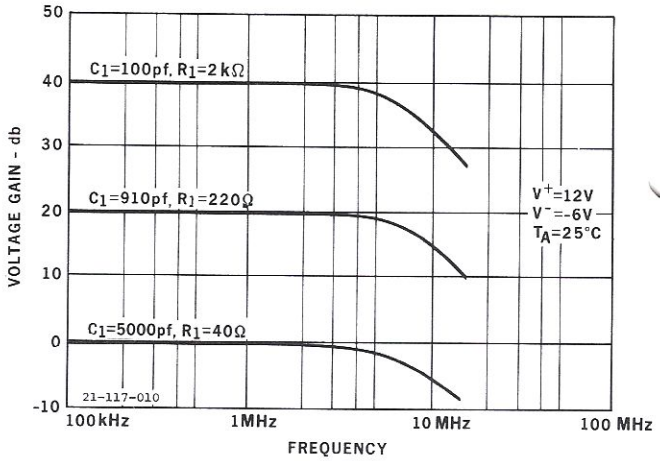
Since variations in the terms R_i, A_{om}, and A_a are correlated to some degree, a compensation network design using individual limits for these parameters would be unnecessarily restrictive. It is better to specify the compensation taking the entire quantity $R_i\frac{A_a}{A_{om}}$ as a single variable. Thus the distribution of this quantity, rather than the distributions of the separate terms, should be used in determining the values of the compensation components. Measurement of a large number of circuits gives the following design equations for the compensation network:

$$R_1 = 20\ (1 + R_f/R_s)\quad \Omega \quad (8)$$

$$C_1 = \frac{0.01}{1 + R_f/R_s}\quad \mu f \quad (9)$$

This compensation will give stable closed-loop gain for the full production distribution of units.

Examples of the closed-loop responses obtained using the recommended frequency compensation are given in Fig. 8. The bandwidth can be expected to vary from 4 MHz to 10 MHz with less than 3 db of peaking, depending upon the particular unit.



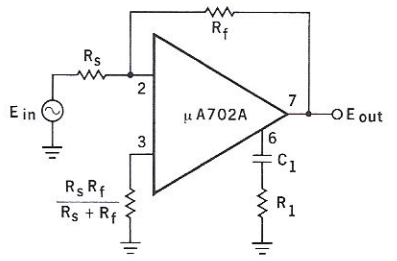


FIG. 8. Frequency response of amplifiers with lag compensation.

The source and feedback resistors should be kept as small as possible for maximum bandwidth. High resistances reduce the bandwidth because of finite amplifier input resistance and stray capacitance shunting the feedback resistor. If maximum bandwidth is not required, additional stability margin can be obtained by reducing R_1 and increasing C_1.

The frequency response of the μA702A is primarily a function of the collector-base capacitances of the integrated transistors, resistance values, and stage gains. Since these parameters are temperature sensitive, the bandwidth will change with temperature. Even though the absolute value of the bandwidth varies, the ratios between the break frequencies stay fairly constant, and therefore the phase margin does not change much. Thus, the circuit will be stable over wide temperature ranges. The change in closed-loop bandwidth with temperature is approximately -0.5%/°C for low source resistances, as shown in Fig. 9. With high source resistances, the bandwidth variation is offset by the change in loop gain due to the temperature coefficient of the amplifier input resistance; hence the relative change in closed-loop bandwidth is much less.

The effect of changes in the negative supply voltage on the bandwidth is shown in Fig. 10. If much variation in voltage is expected, the values of the compensation network components, as given in Eqs. (8) and (9), should be adjusted to give an adequate phase margin for the largest supply voltage anticipated. Dividing R_1 and multiplying C_1 by a factor of 1.5 to 2 will guarantee stability for voltages up to -9 volts. The bandwidth is not significantly affected by the positive supply voltage.

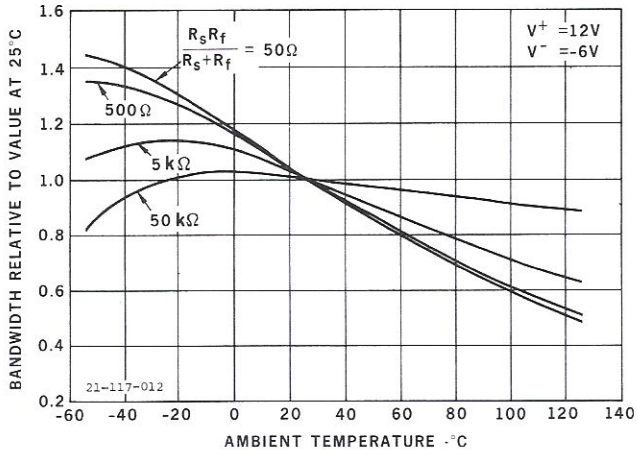


FIG. 9. Effect of temperature on response of a lag-compensated amplifier.

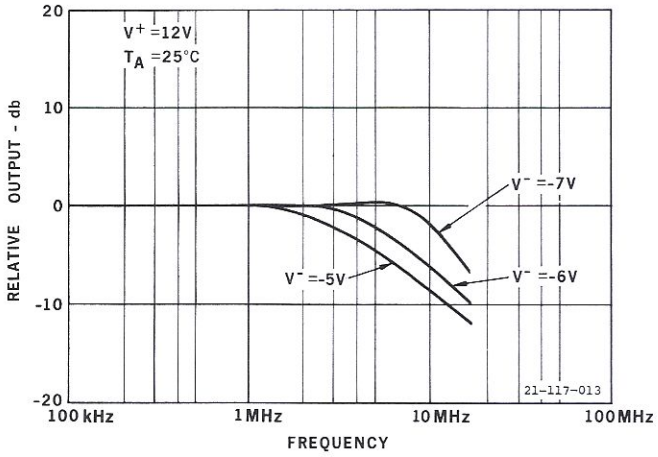
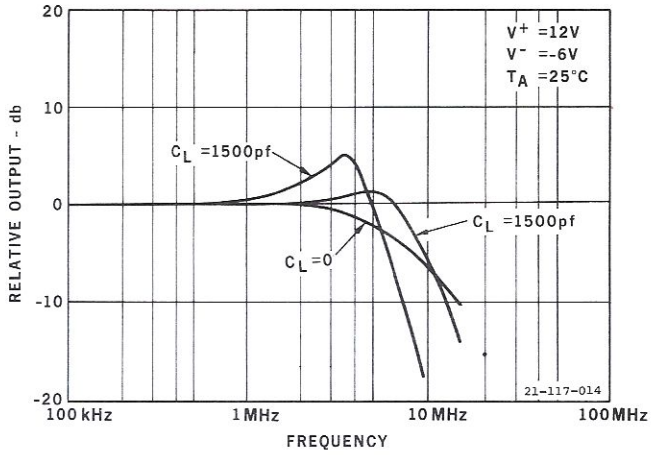


FIG. 10. Effect of negative supply voltage on response of a lag-compensated amplifier.

Excessive capacitive loading at the output of a feedback amplifier may cause peaking and possible instability. The capacitance breaking with the amplifier output resistance produces additional phase shift at the unity loop gain frequency, decreasing the phase margin. Since the compensation network is at the base of the output transistor, it reduces the high-frequency output impedance and makes the circuit less sensitive to capacitive loading. The effect of load capacitance on a unity-gain amplifier is shown in Fig. 11. Selection of the compensation network values for stability with high capacitive loading may be done the same as for supply voltage variations.

Input Lag Compensation

A significant disadvantage of frequency compensating near the output of the amplifier is that the maximum high-frequency output voltage swing is reduced by the compensation network. Since the compensation terminal is at a high-impedance, high-signal level point within the circuit,



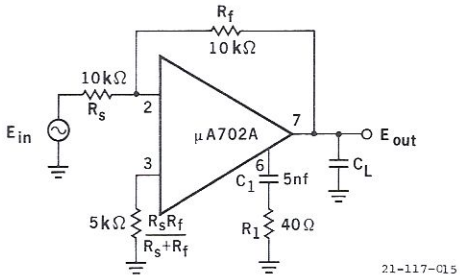


FIG. 11. Effect of load capacitance on response of a unity-gain lag-compensated amplifier.

shunting the load impedance of Q_6 with the compensation network produces a proportional decrease in the peak undistorted voltage swing of the stage. Hence, full output swing can be obtained up to only the compensated open-loop bandwidth of the amplifier.

This limitation may be overcome by placing the compensation network at a low level point in the system. Connection of the network across the input terminals, as in Fig. 12, permits the full output swing capability of the amplifier to be utilized. The attenuation in loop gain from the input compensation network is given by

$$A_c = \left(\frac{2\,\frac{R_s R_f}{R_s + R_f}}{R_2} \right) \left(\frac{R_{in}}{R_{in} + 2\,\frac{R_s R_f}{R_s + R_f}} \right) \qquad (10)$$

for '

$$R_{in} >> R_2 << 2\,\frac{R_s R_f}{R_s + R_f}$$

From Eqs. (3), (5), (6), and (10), the compensation components are

$$R_2 = 2\,R_f\,\frac{A_a}{A_{om}} \qquad (11)$$

$$C_2 = \frac{A_{om}}{4\pi f_o R_f A_a} \qquad (12)$$

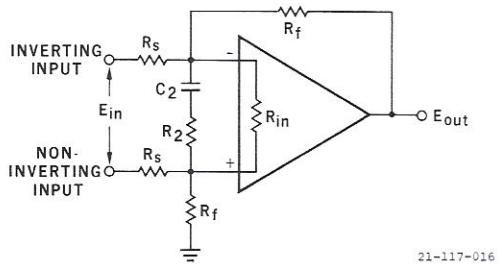


FIG. 12. Amplifier with lag compensation at the input.

These expressions show that input compensation gives another advantage in addition to full output swing capability: the compensation is independent of the amplifier input resistance. This means that changes in input resistance will not affect the closed loop bandwidth. Also, there will be less spread in bandwidth from unit to unit since the variations caused by R_i are eliminated.

The values for the compensation components are found in the same manner as for output lag compensation:

$$R_2 = 5R_f \quad \Omega \qquad (13)$$

$$C_2 = \frac{0.04}{R_f} \quad \mu f \qquad (14)$$

where R_f is in kilohms.

Lead Compensation

The compensation techniques mentioned thus far are applicable to both the μA702 and μA702A amplifiers. The μA702A, however, has an additional compensation point which can be used to put some leading phase shift into the open-loop response and thus increase the useful bandwidth by about two octaves. Connecting a small capacitor across the lead-lag frequency compensation terminals (see Fig. 4) effectively moves the second break point in the open-loop frequency response to a much higher frequency. This gives the frequency response shown in Fig. 13; about 30 db of feedback may be used for a 45-degree minimum phase margin. Thus 10 db more feedback can be obtained than for lag compensation only, and the bandwidth is about four times as large. The optimum value for the lead compensation capacitor is about 50-100 pf.

Closed-loop gains less than 40 db require additional lag compensation. This is best placed at the input, both to reduce the effective input impedance and to maintain a large high-frequency output swing capability. The design equations for the lag compensation components are:

$$R_2 = 20R_f \quad \Omega \qquad (15)$$

$$C_2 = \frac{0.01}{R_f} \quad \mu f \qquad (16)$$

where R_f is in kilohms.

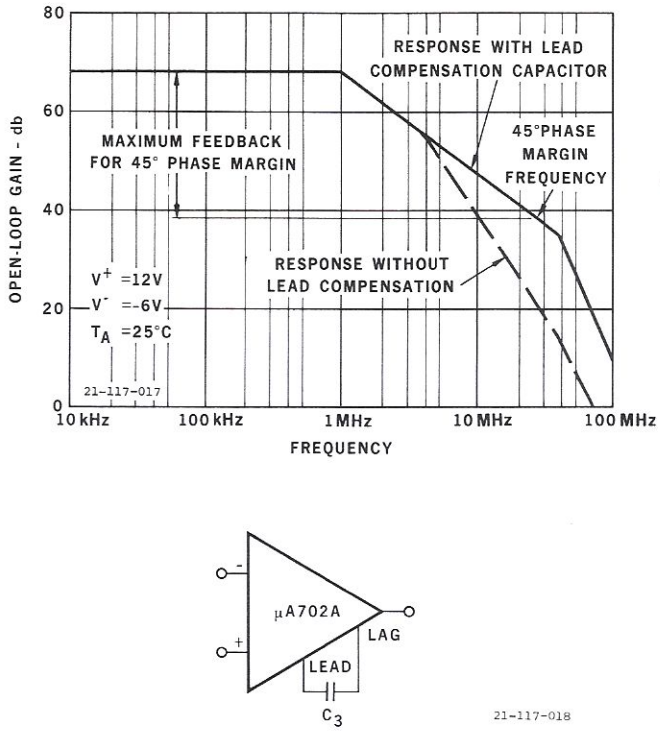
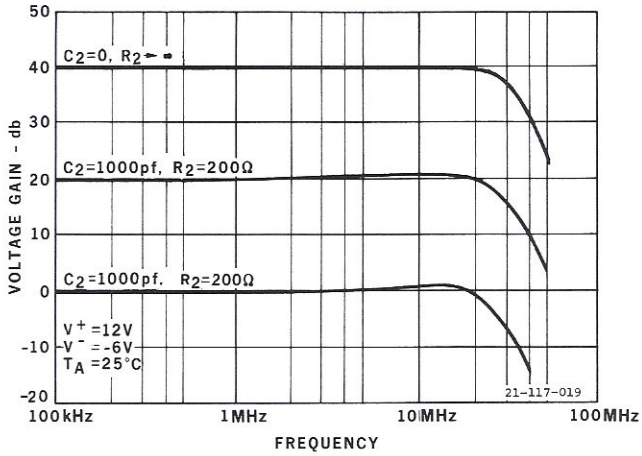


FIG. 13. Open-loop response of μA702A with lead compensation.

Examples of the closed-loop responses obtained using both lead and lag compensation are shown in Fig. 14. The bandwidth will vary from 12 MHz to 40 MHz depending upon the open-loop gain and bandwidth of the particular unit. With the large bandwidths possible using this compensation, it is



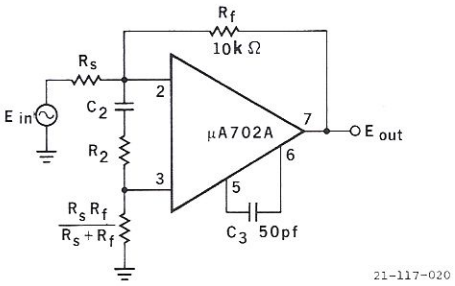


FIG. 14. Frequency response of lead-compensated amplifiers.

extremely important to keep the impedance levels low to avoid undesirable effects due to stray capacitance. Only 0.5 pf across a 30-kΩ feedback resistor, for example, will reduce the bandwidth to 10 MHz.

The relative change of bandwidth with temperature is shown in Fig. 15. The independence of bandwidth from source and input resistances is clearly evident. The effect of negative supply voltage, Fig. 16, is about the same as for output lag compensation.

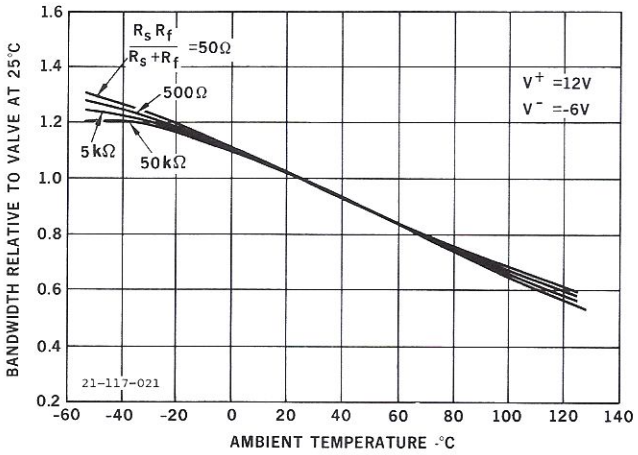


FIG. 15. Effect of temperature on response of a lead-compensated amplifier.

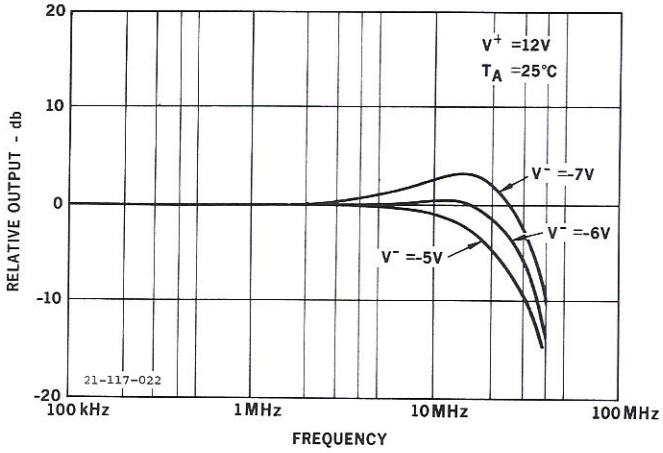
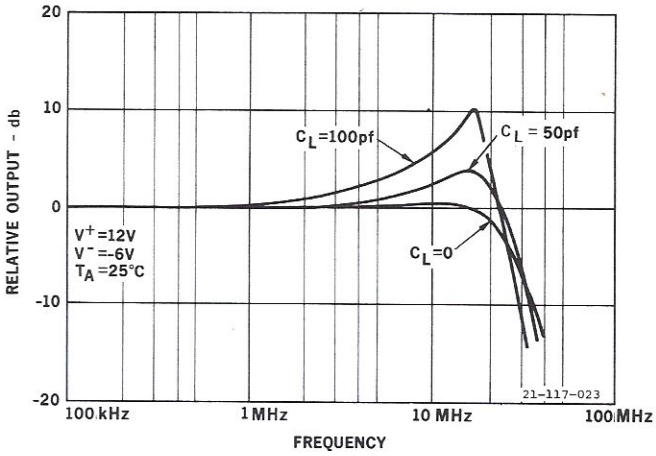


FIG. 16. Effect of negative supply voltage on response of a lead-compensated amplifier.

The circuit is considerably more sensitive to capacitive loading, however, since the bandwidth is so much larger. The peaking due to load capacitance (see Fig. 17) may be reduced by shunting the feedback resistor with a very small capacitor or by increasing the lead capacitor to about 500 pf. This is effective if the load capacitance isn't too large and is relatively constant. For load capacitances greater than 100 pf, it is best to increase the phase margin by dividing R_2 and multiplying C_2 by a factor of 2 or more.



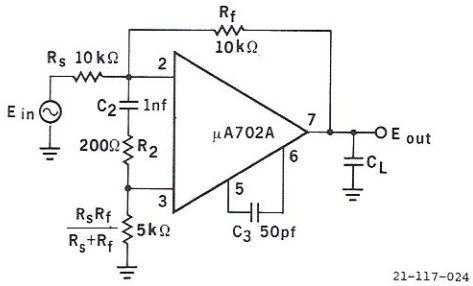


FIG. 17. Effect of load capacitance on response of a unity-gain lead-compensated amplifier.

Other Circuits

The design equations derived for the frequency compensation networks are sufficiently conservative to guarantee stability for the full production distribution of μA702's and μA702A's. For certain specific applications or for a few selected units, it may be advantageous to experimentally determine the optimum compensation for maximum bandwidth. The proper procedure is to measure the loop-gain response with the circuit of Fig. 18, adjusting the compensation for the desired phase margin at the unity loop-gain point. Thus maximum bandwidth can be obtained for given conditions of supply voltages, temperature, loading and other circuit variables.

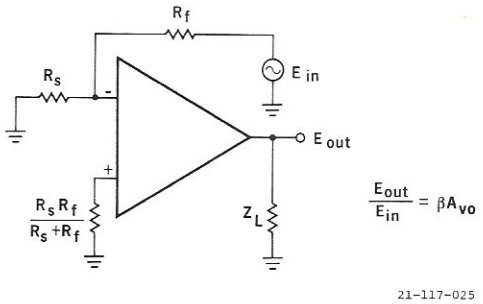


FIG. 18 Circuit for measuring loop gain.

The compensation components for operation at +6, -3 volt supplies are readily calculated from the +12, -6 volt values. The open-loop frequency response and internal impedances are fairly constant with supply voltage; the only important change is that the open-loop voltage gain is about 10 db less. Hence, the proper values for the lag networks may be found by multiplying R and dividing C by a factor of 3. The 50-pf lead compensation capacitor remains the same. Following this procedure gives essentially the same bandwidths and stability margin as for high-voltage supplies.

Special Precautions

Most problems encountered in the use of the μA702A are caused by poor circuit layout and inadequate power-supply bypassing. It should be kept in mind that the amplifier is potentially unstable to almost 100 MHz since it has greater than unity gain below this frequency. Therefore, care should be taken in connecting the amplifier to the outside world to make sure that parasitic oscillations do not occur. This applies to DC circuits as well as to 30-MHz amplifiers.

Of particular importance is power-supply bypassing, as is the case with any high-gain feedback amplifier. Positive feedback through excessive impedance in the power supplies may cause very high frequency internal oscillations in the amplifier. This type of oscillation generally makes itself known by unreasonable DC performance of the amplifier; i.e., the output null may be extremely sensitive to power-supply voltage changes, common-mode inputs, or signal polarity. Bypassing the power supplies right at the amplifier socket with 0.01 to 0.1 μf low-inductance capacitors will eliminate this form of possible instability.

If the circuit layout is not neat, stray capacitance from the output to the inputs may cause excessive peaking in the response or actual oscillations. Even with a neat layout, the unavoidable stray capacitance may cause problems if the source resistances are high. This may be cured by using an inverting amplifier configuration and bypassing the non-inverting input to ground with 50 to 100 pf. This remedy cannot be used if the compensation network is at the input, however, but in this case the network reduces the effective input impedance so that stray capacitance is less critical.

Some difficulty may be experienced when thermal tests are performed on the amplifier. The capacitive coupling between the long leads from the thermal test jig to the external circuit components can make the amplifier hard to stabilize. Under these circumstances it is mandatory to keep the leads as short as possible, with connections to the external circuitry made immediately outside of the temperature chamber. The frequency compensation network should be very conservative, and may be mounted directly on the amplifier socket with the power-supply bypasses, if necessary. The output should be monitored with a high-gain, wideband oscilloscope to make sure the setup is not oscillating before proceeding with the measurements.

CONCLUSION

Circuit stability using integrated operational amplifiers can be obtained using the frequency compensation techniques presented. Specification of the complete compensation network for a given amplifier takes into account the correlation existing between the individual circuit parameters and permits maximum bandwidth to be obtained. Use of the equations and curves given for the μA702A permits stable feedback circuits of practically any configuration to be designed using this amplifier without extensive laboratory measurements and which will be stable for the full production distribution of units.

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